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CERTIFICATE

This certificate is issued in support of an application for Patent registration in a country outside New Zealand pursuant to the Patents Act 1953 and the Regulations thereunder.

I hereby certify that annexed is a true copy of the Provisional Specification as filed on 25 June 2003 with an application for Letters Patent number 526669 made by INDUSTRIAL RESEARCH LIMITED.

Dated 8 July 2004.

Neville Harris

Neville Harris
Commissioner of Patents, Trade Marks and Designs

COMMISSIONER OF PATENTS, TRADEMARKS & DESIGNS
NEW ZEALAND

526669

NEW ZEALAND

PATENTS ACT, 1953

PROVISIONAL SPECIFICATION

NARROWBAND INTERFERENCE SUPPRESSION FOR OFDM SYSTEMS

We, INDUSTRIAL RESEARCH LIMITED, a New Zealand company, of Brooke House, 24 Balfour Road, Parnell, Auckland, New Zealand, do hereby declare this invention to be described in the following statement:

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FIELD OF INVENTION

The invention relates to narrowband interference suppression systems for OFDM communications system and in particular to excision filtering for narrowband interference suppression system for OFDM communication systems.

BACKGROUND

Orthogonal frequency division multiplexing (OFDM) has become the physical layer of choice for many wireless communications systems. An attractive feature of current wireless local area network (WLAN) and wireless metropolitan area network (WMAN) standards based on OFDM is the designed ability to operate in unlicensed spectrum. However, these systems must share spectrum with other unlicensed systems; such as cordless telephones, garage door openers, baby monitors and microwave ovens; which produce narrowband interference in WLAN and WMAN systems. Further, radio non-idealities such as transmitter carrier feedthrough (also known as carrier leakage) also introduce narrowband interference in the form of single-tone carrier residues.

Pilot symbol assisted systems are particularly susceptible to narrowband interference during receiver detection and synchronisation. Pilot symbol assisted systems are also susceptible to narrowband interference on the data transport phase of receiver operation. An interference suppression technique has been proposed to improve the performance of pilot symbol assisted detection and synchronisation in the presence of narrowband interference. However, this technique cannot be applied during data transport as it introduces inter-symbol interference.

Previous proposed interference suppression systems for OFDM include using pre-coding, spread spectrum OFDM, and post-detection receiver techniques involving equalizers. There are many literature reports on narrowband interference suppression techniques for spread spectrum systems, including excision-based methods. However the interference suppression requirements for OFDM differ significantly from the requirements for spread spectrum.

A common model for a received, baseband (low pass equivalent) OFDM symbol, sampled with period T, is

$$r_n = as\left(nT - \tau_S\right)e^{-j\left[2\pi\nu\left(nT-\tau_S\right)+\theta\right]} + \eta\left(nT\right), \tag{1}$$

where a is the flat fading channel amplitude, $s(t)$ is the transmitted signal, n is the sample index, τ_S, ν and θ are, respectively, the time-, frequency- and phase-offsets between transmitter and receiver introduced by a combination of system non-idealities and channel linear distortions, and η is complex additive white Gaussian noise (AWGN) having variance σ_w^2. This model requires a number of assumptions including that the multipath channel is frequency non-selective (flat) and non time-varying (static).

Assuming perfect synchronisation, then, consideration of narrowband interference using the model of equation 1 produces

$$\begin{aligned} r_n &= as\left(nT\right) + be^{-j\left[2\pi\xi nT+\phi\right]} + \eta\left(nT\right) \\ &= as_n + be^{-j\left[2\pi\xi nT+\phi\right]} + \eta_n, \end{aligned} \tag{2}$$

where b, ξ and ϕ are, respectively, the amplitude, frequency and phase of the *demodulated* narrowband interferer. Carrier feedthrough in the transmitter produces an in-band interferer at a frequency equal to the frequency difference between transmitter and receiver local oscillators which, depending on the amount of Doppler shift, will be equal or close to the frequency offset, ν. Typically, the maximum carrier frequency offset is much less than the OFDM sub-carrier spacing and the pilot symbol is designed specifically to be able to resolve this frequency without ambiguity. Any DC offset will occur at narrowband interference frequency, $\xi=0$ and interference from other users of license-free spectrum may occur either singly (e.g. garage door openers, baby monitors, microwave ovens) or in pairs (e.g. cordless telephones) at any in-band frequency. It is noted that, at the time of writing, anecdotal evidence suggests a much higher likelihood of interference in the 2.4 GHz ISM band than in the 5 GHz band. This suggests that the principal immediate application of the invention may be to IEEE 802.11g-compliant WLANs.

At the transmitter, an OFDM symbol is produced such that

$$s_n = w_n \sum_{k=0}^{L-1} d_k e^{j2\pi n \frac{k}{L}}, \tag{3}$$

is the nth of L samples in the OFDM symbol, where d_k is the kth data symbol from some modulation constellation (for example, m-PSK or m-QAM), and w_n is a windowing function which is often simply rectangular ($w_n=1$, $\forall n$). Typically, a K member subset of $\{d_k\}$ is set to zero as spectral blanking, thus L time domain samples represent L-K frequency domain symbols.

At the receiver, the L-point inverse discrete Fourier transform (DFT) of the narrowband interferer has the kth sample

$$\begin{aligned} I_k &= \sum_{n=0}^{L-1} b e^{-j[2\pi\xi nT+\phi]} e^{-j2\pi k\frac{n}{L}} \\ &= b\Psi_k(\xi,\phi), \end{aligned} \tag{4}$$

where

$$\Psi_k(\xi,\phi) = e^{-j\left[\pi(L-1)\left(\frac{k}{L}-\xi T\right)+\phi\right]} \frac{\sin \pi L\left(\frac{k}{L}-\xi T\right)}{\sin \pi\left(\frac{k}{L}-\xi T\right)} \tag{5}$$

is the sampled circular sinc function centred at the interferer frequency. Where the interferer frequency does not coincide exactly with a DFT frequency sample (that is, where $\xi T \neq k/L, k \in \{0 \ldots L-1\}$) then $|I_k| > 0, \forall k$, which is known as *spectral leakage*. Thus, the amount of interference experienced by each data symbol depends on the particular value of the interferer frequency.

It is straight-forward, for any assumed modulation constellation, to use equations (3) and (4) to produce an average BER as a function of signal-to-noise ratio (SNR) and interference-to-noise ratio (INR) over the ensemble of ξT and ϕ. For example, the average BER for BPSK modulated OFDM is given by the marginal distribution

$$\begin{aligned} p_e(\gamma_b,\gamma_i) &= \iiint_{k\xi\phi} p_e(\gamma_b,\gamma_i|k,\xi,\phi)\, p_k(k)\, p_\xi(\xi)\, p_\phi(\phi)\, d\phi\, d\xi\, dk \\ &= \frac{1}{L-K}\sum_{k=0}^{L-K-1} \frac{1}{2\pi}\int_{-\pi}^{\pi} T\int_0^{\frac{1}{T}} p_e(\gamma_b,\gamma_i|k,\xi,\phi)\, d\phi\, d\xi, \end{aligned} \tag{6}$$

where

$$p_e\left(\gamma_b,\gamma_i \middle| k,\xi,\phi\right)=\left[\frac{1}{4}\mathrm{erfc}\left(\sqrt{\gamma_b}+\sqrt{\gamma_i}\,\mathrm{Re}\left\{\Psi_k\left(\xi,\phi\right)\right\}\right)\right.$$
$$\left.+\frac{1}{4}\mathrm{erfc}\left(\sqrt{\gamma_b}-\sqrt{\gamma_i}\,\mathrm{Re}\left\{\Psi_k\left(\xi,\phi\right)\right\}\right)\right] \tag{7}$$

is the BER conditioned on the particular values of k, ξ and ϕ; $\gamma_b = Ad^2/\sigma_W{}^2$ is the SNR per bit for (constant) BPSK bit energy d^2, $\gamma_i = b^2/\sigma_W{}^2$ is the mean INR per bit, σ_W^2 is the noise power per bit, and noting that only L-K frequency bins carry data. Note that A and σ_W^2 are frequency domain duals of the time domain quantities a and σ_w^2 expressed in equations (1) and (2). Note also that the effect of frequency-selective fading additionally can be accounted for by replacing frequency flat A with A_k, such that the summation in equation (6) additionally averages channel gain across frequency.

Expressions similar to equation (6) have been developed previously for evaluating the BER of direct sequence spread spectrum (DSSS) suffering narrowband interference. The principal difference is that the previously developed expressions average SNR and INR across frequency *within* the error function, whereas equation (6) averages SNR and INR across frequency *outside* the error function. This highlights that BER for DSSS depends on *average* narrowband interference power, thus is *insensitive* to particular values of interferer carrier frequency and phase, and so any techniques that reduce *average* SIR will improve BER performance. By contrast, BER for OFDM depends on interference power *per DFT bin*, thus is *sensitive* to particular values of interferer carrier frequency and phase, and the *DFT itself* increases BER by introducing interference to more data-bearing carriers through spectral leakage. Moreover, any interference suppression technique which introduces correlation across frequency bins may *further increase* the BER for OFDM by introducing inter-carrier interference. This shows that interference suppression systems developed for DSSS are not optimised for narrowband interference suppression in OFDM system and in some cases may even increase interference.

Ensemble expressions, such as equation (6), are of limited value in predicting the BER for a given received packet, as there is considerable variation within the ensemble and

the particular values of the interferer frequency, ξT, and interferer phase, ϕ, largely will determine the BER for an individual received packet. This is illustrated in Figures 1A and 1B, which show analytical bit error rates for particular values of interferer frequency, ξT, and interferer phase, ϕ, as well as for the ensemble average (both analytical and simulation) and for interference-free (ideal) BPSK modulated OFDM. Note that the SNR and signal-to-interference ratio (SIR) are both expressed as *average per sample* quantities in Figure 1A and 1B. As can be seen from these Figures the bit error rates are highest for the interferer where the interferer frequency, ξT, is not equal to the DFT frequency sample, *R/L*, and the phase of the interferer is *p/2*. It should be noted that the legend in Figure 1B applies also to Figure 1A. These figures show close agreement of ensemble averages produced analytically and by computer simulation. Note also the variation within the ensemble, indicated both by analytical BER curves for particular values within the ensemble and by the 10th and 90th percentile limits on the computer simulation BER curves (shown as error bars).

Inspection of equations (6) and (5) indicates that the use of a windowing function in equation (3) can significantly *decrease* the effect of narrowband interference on most frequency domain data symbols for some values of interferer frequency ξT. However, in the case where $\xi T \cong {}^{k}\!/_{L}, k \in \{0 \ldots L-1\}$, the amount of spectral leakage is minimal, and any windowing function will introduce spectral spreading close to the interferer frequency which will significantly *increase* the effect of interference on neighbouring frequency domain data symbols. A windowing function may also produce inter-carrier interference, which will lead to bit errors. Thus, the particular value of interferer frequency ξT not only determines the BER experienced by the receiver but also determines the effectiveness of time-domain windowing as an interference-suppression technique. Unless the interferer frequency is known in advance, which will not be the case in real world systems, the effectiveness of time-domain windowing cannot be determined in advance. Thus time-domain windowing systems are not useful as they cannot be guaranteed to reduce interference and in some cases increase interference.

It has been shown that application of the N-LMS algorithm provided significant interference suppression to pilot symbol based detection and synchronisation

algorithms. However, it has also been shown that the N-LMS algorithm introduced inter-symbol interference (ISI) as a result of producing the reference signal as a delayed version of the input signal. There are a variety of techniques which can be used to prevent ISI being introduced by the N-LMS algorithm. Each of these techniques, however, effectively require that subtractive interference cancellation operate using initial interference parameter estimates which have aged by as much as the maximum duration of a data packet – about 5.5 ms (10^5 samples) for IEEE 802.11a. Figure 2 shows the effectiveness of subtractive interference cancellation based on aged parameter estimates where the initial estimation accuracy achieved the Crámer-Rao lower bound. This result leads to the conclusion that subtractive interference cancellation algorithms are not suitable for application to data transport functions.

In Figure 2 the effect of aging parameter estimates on subtractive interference cancellation is shown. In this Figure L is the number of samples used to produce the initial estimates of interferer amplitude, carrier frequency and phase. The dotted lines shown the results where no parameter estimates are updated, whereas the solid lines show the results where only the carrier frequency estimate is *not* updated. The dashed lines show the (uncancelled) interferer magnitudes.

SUMMARY OF INVENTION

It is the object of the present invention to provide an improved system and method for interference suppression in OFDM communications systems or to at least provide the public with a useful choice.

In broad terms in one aspect the invention comprises a method for suppressing narrowband interference in OFDM receivers including the steps of acquiring a sample of received data, estimating parameters of each of a number of narrowband interferers from the acquired sample of data, forming an excision filter using the estimated parameters and inserting the excision filter into an OFDM receiver.

Preferably the filter is inserted into the OFDM receiver prior to a discrete Fourier transform.

Preferably the estimated parameters of the narrowband interferers include demodulated carrier frequency, magnitude and phase.

Preferably the step of estimating the number of narrowband interferers includes the steps of performing a forward DFT on the samples, and performing a periodogram search on the output of the DFT to identify peaks in the periodogram where the number of peaks in the periodogram corresponds to the number of interferers.

Preferably the step of estimating parameters of the narrowband interferers includes estimating the frequency of an interferer as the frequency of a peak on the corresponding periodogram, estimating the magnitude of the interferer as the amplitude of the corresponding periodogram peak, and estimating the phase of the interferer as the phase of the corresponding periodogram peak.

Preferably the narrowband interferer parameter estimates of each narrowband interferer are used to initialise a digital phase locked loop.

Preferably the method for suppression narrowband interference includes the step of receiving an indication of a start of packet when a data packet is received by the OFDM receiver.

Preferably the phase locked loops are updated with each incoming sample until either a counter expires or an OFDM packet is detected. The phase locked loops are used to estimate the carrier frequency of the narrowband interferers. Preferably the phase locked loops are digital phase locked loops. Preferably one phase locked loop is used for each interferer.

Preferably the current narrowband interferer carrier frequency estimates from the phase locked loops that have achieved "lock" are used to initialise an excision filter when an OFDM packet is detected.

Preferably the excision filter has impulse response duration less than the OFDM guard interval.

In broad terms in a further aspect the invention comprises an OFDM receiver including a front end arranged to receive data, a data sampler arranged to acquire a sample of received data, a parameter estimator that estimates parameters of each narrowband interferer in the sample of received data and an excision filter that uses the estimated parameters of each narrowband interferer to reduce noise from the narrowband interferers.

Preferably the filter is inserted into the OFDM receiver prior to a discrete Fourier transform.

Preferably the parameter estimator estimates the demodulated carrier frequency, magnitude and phase of the narrowband interferers.

Preferably the parameter estimator is also arranged to perform a forward DFT on the samples, and perform a periodogram search on the output of the DFT to identify peaks in the periodogram where the number of peaks in the periodogram corresponds to the number of interferers.

Preferably the parameter estimator is also arranged to estimate the frequency of an interferer as the frequency of a peak on the corresponding periodogram, estimate the magnitude of the interferer as the amplitude of the corresponding periodogram peak, and estimate the phase of the interferer as the phase of the corresponding periodogram peak.

Preferably the narrowband interferer parameter estimates of each narrowband interferer from the parameter estimator are used to initialise a digital phase locked loop.

Preferably OFDM receiver is further arranged to provide an estimate of the start of an OFDM data packet to the parameter estimator.

Preferably the parameter estimator includes phase locked loops that are updated with each incoming sample until either a counter expires or an OFDM packet is detected. The phase locked loops are used to estimate the carrier frequency of the narrowband interferers. Preferably the phase locked loops are digital phase locked loops. Preferably one phase locked loop is used for each interferer.

Preferably the current narrowband interferer carrier frequency estimates from the phase locked loops that have achieved "lock" are used to initialise an excision filter when an OFDM packet is detected.

Preferably the excision filter has impulse response duration less than the OFDM guard interval.

BRIEF DESCRIPTION OF DRAWINGS

The invention will be further described by way of example only and without intending to be limiting with reference to the following drawings, wherein:

Figure 1A shows the bit error rate performance of BPSK modulated OFDM with a single interferer and signal to interference ratio (SIR) of -10 dB;

Figure 1B shows the bit error rate performance of BPSK modulated OFDM with a single interferer and SIR of 10 dB;

Figure 2A shows the effect of ageing parameter estimates on interference cancellation with interference to noise ratio (INR) of 20 dB and using 64 samples to produce estimates of the interferer parameters;

Figure 2B shows the effectiveness of interference cancellation with INR of 20 dB and using 1000 samples to produce estimates of the interferer parameters;

Figure 3A is a flowchart showing one technique for interference suppression of the invention;

Figure 3B is a block diagram of an OFDM receiver including the interference suppression system of the invention;

Figure 4A shows a simulation of interferer carrier frequency estimation where the INR is 7.5 dB;

Figure 4B shows a simulation of interferer carrier frequency estimation where the INR is 6.6 dB;

Figure 4C shows phase lock loop indication for the interferers of Figures 4A and 4B;

Figure 5A shows a prototype excision filter frequency response;

Figure 5B shows an example of a two notch excision filter;

Figure 6A shows an example of a received signal with the narrowband interference suppression system in place;

Figure 6B shows the smoothed spectra of a first OFDM data block after narrowband interference is suppressed;

Figure 7A shows a received signal constellation including signal, narrowband interference and noise;

Figure 7B shows the received signal constellation after filtering to remove narrowband interference;

Figure 7C shows the received signal constellation when no narrowband interference is present;

Figure 8A is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SIR of -10 dB;

Figure 8B is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SIR of 0 dB;

Figure 8C is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SIR of 10 dB;

Figure 9A is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SNR of 6 dB;

Figure 9B is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SNR of 12 dB;

Figure 9C is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SNR of 18 dB;

Figure 10A is a simulation of bit error rates for 64-QAM modulated OFDM with two narrowband interferers and SIR of 0 dB; and

Figure 10B is a simulation of bit error rates for BPSK modulated OFDM with two narrowband interferers and SIR of -15 dB.

DETAILED DESCRIPTION

Excision-based methods of interference cancellation require estimation of the carrier frequency only for each narrowband interferer. For an excision filter having a frequency response $H(f)$, the post-excision conditional BER for BPSK modulated OFDM is produced by modifying equation (7) to obtain

$$p_e(\gamma_b,\gamma_i|k,\xi,\phi)=\left[\frac{1}{4}\operatorname{erfc}\left(\sqrt{\gamma_b}+\frac{H(\xi)}{H\left(\frac{k}{T}\right)}\sqrt{\gamma_i}\operatorname{Re}\{\Psi_k(\xi,\phi)\}\right)\right.$$
$$\left.+\frac{1}{4}\operatorname{erfc}\left(\sqrt{\gamma_b}-\frac{H(\xi)}{H\left(\frac{k}{T}\right)}\sqrt{\gamma_i}\operatorname{Re}\{\Psi_k(\xi,\phi)\}\right)\right], \quad (8)$$

which may be inserted into equation (6) to produce the marginal distribution representing the ensemble BER. Each OFDM sub-channel for which the frequency response of the OFDM data is greater than the frequency response of the narrowband interferer, $H\left(\frac{k}{T}\right)>H(\xi)$, will experience an improved BER compared to the unfiltered equivalent. In particular when the frequency response of the narrowband filter is zero, $H(\xi)=0$, equation (8) reduces to the ideal BER for BPSK. Note that equation (8) is valid only for excision filters having an impulse response duration *less* than the OFDM guard interval. Where this is not the case, the excision filter will introduce some degree of inter-symbol interference.

The main challenges for excision filtering are, firstly, estimation of the interferer carrier frequency (such that the notch of the excision filter is correctly positioned in frequency) and, secondly, design and implementation of an effective and efficient notch filter. To account explicitly for the effect of carrier frequency estimation error on equation (8), the frequency response of the narrowband interferer, $H(\xi)$, can be modified to produce the excision filter frequency response conditional on the estimation error $H(\xi|\varepsilon_\xi)$, where $\varepsilon_\xi=\xi-\hat{\xi}$ is the particular value of estimation error and $\hat{\xi}$ is the estimated frequency of

the narrowband interferer. The marginal distribution of equation (6) then must be integrated additionally over the estimation error distribution $p_\varepsilon\left(\varepsilon_\xi\right)d\varepsilon_\xi$ to obtain the ensemble BER. Note that excision filtering does not require the estimation of interferer power, so any time-variation introduced by the fading channel will not reduce the efficacy of excision filtering. Also, provided that the filter notch width is sufficient, excision filtering is insensitive to some time variation in interferer frequency, due to either modulation of the interferer or time variation in the channel.

For a single interferer, the carrier frequency can be estimated directly from the Wiener solution provided by an N-LMS filter. Such an N-LMS filter may be used for pilot symbol interference cancellation. It is straight-forward to show that, for a single interferer defined in equation (2), a three tap N-LMS filter will estimate the Wiener filter coefficients

$$\mathbf{w} = \frac{1}{\det \mathbf{R}} \begin{bmatrix} 2b^2\sigma_w^{\ 2} + \sigma_w^{\ 4} \\ -b^2\sigma_w^{\ 2} e^{-j2\pi\xi T} \\ -b^2\sigma_w^{\ 2} e^{-j4\pi\xi T} \end{bmatrix}, \tag{9}$$

where $\det \mathbf{R}$ is the (real) determinant of the input correlation matrix $\mathbf{R}$. This allows direct estimation of the frequency of the demodulated narrowband interferer, ξ (and the amplitude of the demodulated narrowband interferer, b).

The analogous result for two interferers, having amplitudes b_1 and b_2 and demodulated carrier frequencies ξ_1 and ξ_2, is

$$\mathbf{w} = \frac{1}{\det \mathbf{R}} \begin{bmatrix} 2b_1^{\ 2}b_2^{\ 2}C_1 + 2b_1^{\ 2}\sigma_w^{\ 2} + 2b_2^{\ 2}\sigma_w^{\ 2} + \sigma_w^{\ 4} \\ 2b_1^{\ 2}b_2^{\ 2}C_2 e^{-j\pi(\xi_1+\xi_2)T} - \sigma_w^{\ 2}\left(b_1^{\ 2}e^{-j2\pi\xi_1 T} + b_2^{\ 2}e^{-j2\pi\xi_2 T}\right) \\ 2b_1^{\ 2}b_2^{\ 2}C_1 e^{-j2\pi(\xi_1+\xi_2)T} - \sigma_w^{\ 2}\left(b_1^{\ 2}e^{-j4\pi\xi_1 T} + b_2^{\ 2}e^{-j4\pi\xi_2 T}\right) \end{bmatrix} \tag{10}$$

where $C_1 = 1-\cos\left(2\pi\left(\xi_1-\xi_2\right)T\right)$ and $C_2 = \cos\left(3\pi\left(\xi_1-\xi_2\right)T\right) - \cos\left(\pi\left(\xi_1-\xi_2\right)T\right)$ for convenience. This does not allow direct estimation of the frequencies of the narrowband interferers ξ_1 and ξ_2.

For more than two interferers, the Wiener solution is more complicated, does not cancel more than two interferers using a three tap filter, and does not allow the interferer parameters to be estimated directly. (For no interferers, the Wiener solution is $\mathbf{w} = (1/\det \mathbf{R})\left[\sigma_w^4 \quad 0 \quad 0\right]^T$, and the N-LMS filter coefficients will be an estimate of this.)

These results are problematic, since they require knowledge of the number of interferers and the number of interferers is not known *a priori*. Thus, it is not possible to use the N-LMS filter coefficients alone to produce an excision filter which will robustly suppress an indeterminate number of interferers.

A technique which overcomes many of the problems described above is shown as a flow-chart in Figure 3A. The narrowband interference suppression system for the data transport phase can run in parallel with a technique for interference suppression during pilot symbol assisted detection and synchronisation. The narrowband interference suppression system of the invention relies on estimating interference carrier frequency(s) during the signal-free period between data packets – this reduces the impact of the particular values of SIR and interferer carrier frequency on the algorithm performance. The estimated carrier frequencies are used to specify the excision filter applied to the received signal after detection. The narrowband interference suppression system of the invention is particularly suited for interference suppression during the data transport phase of the data reception. However this technique can also be used during packet detection and synchronisation.

Figure 3 shows the steps of one embodiment of the excision-based narrowband interference suppression technique. When detection is started in step 1 an initial block of L samples of interference plus noise (no OFDM signal) is collected. The number of samples, L, is the block size of the available DFT software/firmware. In preferred embodiments L is typically the OFDM block size.

After the block of L is collected in step 2 maximum likelihood parameters estimates for the desired narrowband interferers are performed using the forward DFT on the block of

L samples. After taking a DFT of the data block an assumption is made of the number of narrowband interferers present. Assuming that M interferers are anticipated as being present, then a periodogram search is employed to identify the M largest periodogram peaks from the DFT output. To avoid false detection of interferers due to spectral leakage, periodogram bins either side of each identified peak are not considered in subsequent searches. The assumption of the number of interferers present is based on the maximum number of interferers that could be present. For example, if interference is caused by a cordless telephone there could be two interferers present but it is less likely that more than two interferers will be present. There is a performance/complexity trade-off in the number of interferers estimated to be present. Estimating a high number of interferers will give better performance but will also increase the complexity of the system compared to estimating a lower number of interferers.

After the periodogram search has identified the amplitude and phase of each identified periodogram peak are the estimated interferer magnitude and phase, respectively as shown in step 3. The frequency corresponding to the peak periodogram location is the estimated interferer demodulated carrier frequency.

In step 4 the M parameter estimate vectors are used to initialise M digital phase locked loops (PLLs). At this point a time-out counter is set.

In step 5 the equation is asked whether a packet has been detected. In the embodiment packet detection is performed by another function (shown as step 12). If no packet has been detected the 'no' arrow is followed from step 5 to step 10. If a packet has been detected the 'yes' arrow is followed from step 5 to step 6 and packet reception commences.

In step 10 the question is asked has the time-out counter expired. The time-out counter is used to ensure that the interferer estimates are current. For example, if a new interferer appears after the block of L samples is taken in step 1 it will not be compensated for unless periodically the process repeats by using the latest block of L samples to estimate the narrowband interferer parameters. In this way the techniques

should track the M strongest narrowband interferers. If the time-out counter has not expired the 'no' arrow is followed to step 11. If the time-out counter has expired the 'yes' arrow is followed to step 2 and the latest block of L samples is used to estimate the narrowband interferers.

When the 'no' arrow is followed to step 11 the M PLLs are innovated and the time-out counter is incremented. The arrow is then followed to step 5. In this way the M PLLs are innovated with each incoming sample until *either* an OFDM packet is detected by a parallel process *or* the time-out counter expires. Each PLL locks onto an interferer to provide the carrier frequency of the interferer. If a PLL lock is not achieved then it can be assumed that no interferer is present for that PLL. The number of locks achieved indicates the number of interferers present. For example if two PLLs are initialised and innovated and only one achieves a lock it can be assumed that only one interferer is present. When this occurs only one excision filter is produced as the second interferer (if present) is deemed to have negligible impact on the BER.

When a packet is detected the 'yes' arrow is followed from step 5 to step 6. If a packet has been detected, then the current M interferer parameter estimates are used to initialise the excision filter in step 7. Importantly, each PLL provides a *lock* indication (for example $I_{lock} = \mathrm{Im}(\varepsilon_{PLL})/\mathrm{Re}(\varepsilon_{PLL})$, where $\varepsilon_{PLL} = r_n e^{j[2\pi\hat{\xi}nT+\hat{\phi}]}$ is the PLL error for the received signal, noise and narrowband interference, r_n, defined by equation (2) with the transmitted signal s_n=0, approaches zero when phase lock has been achieved) which allows a genuine interferer to be distinguished from thermal noise. Only interferers for which PLL lock is achieved are excised from the OFDM packet.

During data reception, interferers are removed using one or more notch filter(s) centred at the estimated interferer demodulated carrier frequency shown in step 8. This filter is inserted at the input to the OFDM digital receiver *prior* to the forward DFT to prevent spectral leakage.

Figure 3B shows an OFDM baseband receiver comprising four modules. Module 201 contains an A/D converter driver 21, timer 22 and signal conditioner 23; module 202

contains a packet detector 25, frame timer 26, narrowband interference suppression module 24 and first stage receiver 27; module 203 contains a second stage receiver 28; and module 204 contains a decoder 23.

The data received by RF block 20 which is arranged to shift the data back to baseband. The RF block of the receiver may include a low noise amplifier, bandpass filter, quadrature demodulator and frequency down-converter. The baseband data is then sampled by analogue to digital (A/D) converter 21. This converts the received data from an analogue signal to a digital signal. Ideally the A/D converter samples the received data at greater than the nominal bit rate. The sampled signal then passes through signal conditioner 23, that compensates for some of the channel and noise induced distortions. A further purpose of the signal conditioner 23 is to digitally low pass filter the baseband signal to remove out-of-band noise. The data may also be sampled to the nominal bit rate.

Packet detector 25 and frame timing block 26 search for the start of a packet. Packet detector 25 may also provide narrowband interference suppression when the packet is detected. Narrowband interference suppression block 24 applies narrowband interference suppression during the data transport phase of packet reception. In the preferred embodiment this block implements the algorithm of Figure 3A.

Frame timing module 26 may be further arranged to provide a start of packet estimate to timer 22.

After the start of a packet has been detected by packet detect block 25 the packet is passed through first stage receiver 27. This receiver may estimate and compensate for frequency and phase errors in the received data. The first stage receiver also includes a Fourier transform operator that transforms the data from time domain data to frequency domain data.

The data is then passed into the second stage receiver 28. The second stage receiver commences operation on frame detection. Its functions are, initially, to estimate the time varying sub-sample time offset and, throughout the remainder of the frame, to apply symbol timing error correction. The second stage receiver may also be used to update estimates of other time-varying parameters of the received data. The second stage receiver includes a data decision block that makes hard decisions on each data bit (symbol) prior to error detection and correction.

The second stage receiver may include a demodulator. Acting together with the decision process, the demodulator converts the data back from a modulation scheme, such as QPSK or 64 QAM, to binary data. After hard decisions have been made on the data the data streams are converted back from parallel to serial data.

Following the second stage receiver 28 is decoder 29. The decoder decodes the coded data and performs error corrections and/or detection up to the limit of the decoder. The decoder is matched to an encoder in the corresponding OFDM transmitter. For example if the encoder is a Reed-Solomon encoder then the decoder will be a Reed-Solomon decoder. Following decoding of the data the data is then passed to the electronic equipment attached to the receiver as data sink 30. The basic elements of an OFDM receiver are well known and will not be discussed in more detail.

At the completion of OFDM packet detection, the receiver returns to its original state, described by step 1 of Figure 3A in which new interferers are searched for prior to the reception of the next in-coming OFDM packet.

There are a number of design issues raised by the interference suppression technique, principally concerning the PLLs and the excision filter, and there is also a design trade-off between the time-out threshold and the PLL filter bandwidth.

PLL design parameters depend on a number of system-specifics, such as sub-carrier spacing and inter-packet arrival time as well as interferer characteristics. The PLL filter

bandwidth should be set to be no less than the OFDM sub-carrier spacing, as this is the quantisation of the initial, DFT-based interferer carrier frequency estimate.

An example of phase lock loop operation from the simulation implementation used in this paper is shown in Figures 4A to 4C. Failure of at least one PLL to achieve lock typically occurred where one interferer was of significantly lower power than the other. For example when one interferer has so low power that it could not be distinguished from background noise. In these cases the low power interferer will have negligible impact on the BER. In Figures 4A to 4C the initial estimates are made using a 64-point DFT. These initial values are shown by heavy solid lines. Figure 4A shows the frequency estimate for a first narrowband interferer as well as the actual interferer frequency. As can be seen the frequency estimate is close to the interferer frequency. In this Figure the interference to noise ration is 7.5 dB. Figure 4B shows the frequency estimate for a second narrowband interferer. In this Figure the interference to noise ratio is 6.6 dB. As can be seen the frequency estimate is close to the interferer frequency. Figure 4C shows the operation of two phase lock loops, one for each interferer. After the initial estimates subsequent estimate innovations are made using two digital phase locked loops initialised to the DFT-based estimates. A PLL 'lock' is said to have been achieved when the lock magnitude is less than 0.5 – the lock values shown in Figure 4C correspond to the estimates shown in Figures 4A and 4B.

Key elements of the excision filter design include the 3 dB filter bandwidth, which should be set to being equal to the OFDM sub-carrier spacing. This is a trade-off between competing requirements, first, to allow for estimation errors and time variation in the interferer and, secondly, to minimise the impact of the excision filter(s) on data bearing sub-carriers which are unaffected by interference. An efficient implementation for multiple interferes is to design a single, high-pass, excision filter, and then to (complex) frequency-shift the received signal by each estimated interferer demodulated carrier frequency (negated) prior to filtering. The frequency response of the prototype filter is shown in Figures 5A and 5B.

Figure 5A shows a one-notch excision filter frequency response and Figure 5B shows a two-notch excision filter frequency response. In both of these Figures the abscissa scales are normalised to the OFDM frequency bin number and the location of the OFDM data-bearing sub-carriers are indicated by heavy dots. In Figure 5B the notch filter is designed to suppress two interferers with estimated (normalised) carrier frequencies at -2.76 and 13.5. This interpolated FIR filter comprised two sections of 31 taps and 21 taps, respectively, to achieve an overall impulse response equivalent to that of a 165 tap conventional FIR filter.

The efficacy of the proposed technique is established by computer simulation. The OFDM system simulated is baseband-equivalent to uncoded IEEE 802.11a, assuming perfect detection and synchronisation. The receiver estimated carrier frequencies for two interferers, thus two PLLs were employed. For each simulation, two narrowband interferers were approximated using stationary complex cisoids having amplitudes $\{b_1, b_2\} \doteq U[0,1]$ such that the average signal-to-interference ratio per interferer is $SIR = S \big/ \sqrt{\tfrac{1}{2}\left(b_1^{\,2} + b_2^{\,2}\right)}$ for RMS signal magnitude S, demodulated carrier frequencies $\{\xi_1, \xi_2\} \doteq U[-B/2, B/2]$ for OFDM passband 3 dB bandwidth B, and phases $\{\phi_1, \phi_2\} \doteq U[-\pi, \pi]$. For each transmitted packet, a signal-free training period of 1000 samples (50μs for IEEE 802.11a) is provided to allow the PLLs to estimate the interferer carrier frequencies. Figures 4A to 4C show this number of samples is sufficient to achieve a phase lock in most cases. After 1000 samples, the excision filter is enabled in the receiver and the OFDM packet is transmitted. Each OFDM packet comprised 10,000 bits, to enable bit error rates down to 1×10^{-4} to be measured the results presented have an error floor at 1×10^{-4}.

Figure 6A shows an example received packet magnitude, truncated to 3000 samples from the start of simulation. This Figure shows the 1000 sample signal-free training period, the effect of the excision filter on the signal magnitude and the start of the OFDM packet. Two narrowband interferers are present with SIR of 10 dB and SNR of 30 dB. The interpolated spectrum of the first data block (first 64 samples after the training symbol and signal block) from the same example is shown in Figure 6B,

compared to the interference-free and unfiltered spectra for the same data block. The effect of the excision filter on the two interferers (at bin numbers 2.2 and 50.7, respectively) can be seen in this example. In Figure 6B three spectra are shown: the dashed line shows the data symbol with noise and interference; the dotted line shows the data symbol the noise and where the narrowband interference is suppressed using the method of the invention; and the solid line shows the data symbol with noise and no narrowband interference.

Signal constellations for a complete OFDM packet are shown in Figures 7A to 7C. Figure 7A shows the signal constellation of the received signal plus interference and noise of Figure 6A. Figure 7B shows the signal constellation of the received signal of Figure 6A after interference is excised using the method of the invention, and Figure 7C shows the signal constellation of the signal of Figure 6A when no interference is present. The effects of interference and interference suppression on this example constellation can be seen. The accuracy of the simulation BER results can be verified by comparison with analytical BER results, as shown in Figure 1.

Bit error rate curves were produced by simulation, where each point on a curve is the median of 1000 simulations. This allows the effect of variation within the ensemble of interferer relative amplitudes, carrier frequencies and carrier phases to be indicated by error bars on the 10^{th} and 90^{th} percentiles. Figures 8A to 8C show bit error rate curves for signal-to-interference ratios of –10 dB, 0 dB and 10 dB respectively for the same OFDM packets with two unsuppressed narrowband interferes, with excised interferers, and with no interferers. The interferer-free curves closely follow those for conventional BPSK in AWGN.

Figure 8A shows simulation rates of BPSK modulates OFDM with two narrowband interferes and SIR of 10 dB. The dashed-dotted line shows the signal plus interference and noise, the dashed line shows the filtered signal plus interference and noise where the interference has been excised using the method of the invention and the solid line shows the signal with no narrowband interference. The error bars show the 10^{th} and 09^{th} percentiles of the BER over the ensemble interferer relative powers, carrier frequencies,

carrier phases and AWGN. The same legend applies to Figures 8B and 8C. It can be seen from Figures 8A to 8C that the effects of narrowband interference are severe. Interference excision can be seen to significantly improve the bit error rate, by reducing the median "error floor" to about 1×10^{-3}. However, the variation within the ensemble also can be seen to be greater for the interference excision results. This is due to two factors: firstly, the number of data bits per OFDM block affected by excision depend heavily on the particular values of interferer carrier frequency and, secondly, at least one PLL failed to "lock" for at least 15% of all simulations thus increasing the variability of BER results.

Interference excision was found to be effective across a wide range of SIRs, as shown in Figures 9A to 9C. Figures 9A to 9C show bit error rates for BPSK modulated OFDM with two narrowband interferers. Each of the Figures shows three curves being the ensemble median BERs for the same data packets where: the dashed-dotted line represents signal plus noise and narrowband interference; the dashed line represents the signal plus noise where the narrowband interference has been excised using the method of the invention; and the solid line represents the signal plus noise where no narrowband interference is present. These Figures show that OFDM systems employing interference excision as described here exhibit acceptable median BERs of about 1×10^{-3} for SIRs down to –30 dB. However it should be noted also that the variation within the ensemble increases as the SIR decreases, indicating that interference excision is more likely to leave a packet severely errored as the SIR decreases. Figures 9A to 9C also show that the interference excision ceases to improve BER performance when the SIR exceeds about 15 dB. This effect occurs at an SIR of about 15 dB for two reasons: firstly, narrowband interference causes increasingly fewer bit errors at higher SIRs and, secondly, at a given SIR, for lower SNRs, the low INR makes interferer carrier frequency estimation less reliable.

A limitation of interference excision is that it provides less benefit with QAM modulations in comparison with pure phase modulations. Figures 10A and 10B show BER curves for 64-QAM and BPSK modulated OFDM respectively, where the SIR and SNR have been selected so that the interferer plus noise and interferer-free results are

similar between systems. Figures 10A and 10B each include three curves. The dash-dotted curve represents the data packet plus noise and narrowband interference, the dashed line represents signal plus noise where the interference has been excised using the method of the invention, and the solid line represents signal plus noise where no narrowband interference is present. It can be seen that interference excision provides about one order of magnitude less benefit in BER improvement for the 64 QAM system in comparison with the BPSK system. This is due to the excision filter attenuating (or amplifying) to some extent several OFDM data bins around each interferer. This causes few bit errors in BPSK modulated OFDM as phase modulations are robust to amplitude variations. However, QAM systems are very sensitive to amplitude variations, particularly for large constellations, as signal amplitude is part of the data representation. Although this effect could be mitigated somewhat by equalization, use of a more robust modulation seems prudent in an interference-limited environment.

An interference suppression technique of the invention based on excision filtering has been described and shown by computer simulation to produce a significant improvement in the bit error rate of BPSK modulated OFDM compared to no interference suppression. Using this technique, which works within existing OFDM-based standards, it has been shown that acceptable ensemble bit error rates of about 1×10^{-3} are obtainable for signal-to-interference ratios as low as –30 dB. Excision filtering is effective for OFDM, because it significantly reduces the major error-producing effect – spectral leakage – by filtering the interference before the discrete Fourier transform. The interference suppression technique of the invention has particular application for the data transport phase of receiver operation rather than the detection and synchronisation phase (although it can be used during packet detection and synchronisation). The technique of the invention can be used in combination with a technique for narrowband interference detection and suppression during the detection and synchronisation phases of data reception.

The foregoing describes the invention including preferred forms thereof. Alterations and modifications as will be obvious to those skilled in the art are intended to be incorporated in the scope herein.

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By their authorised agents
A. J. PARK
Per

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25 JUN 2003

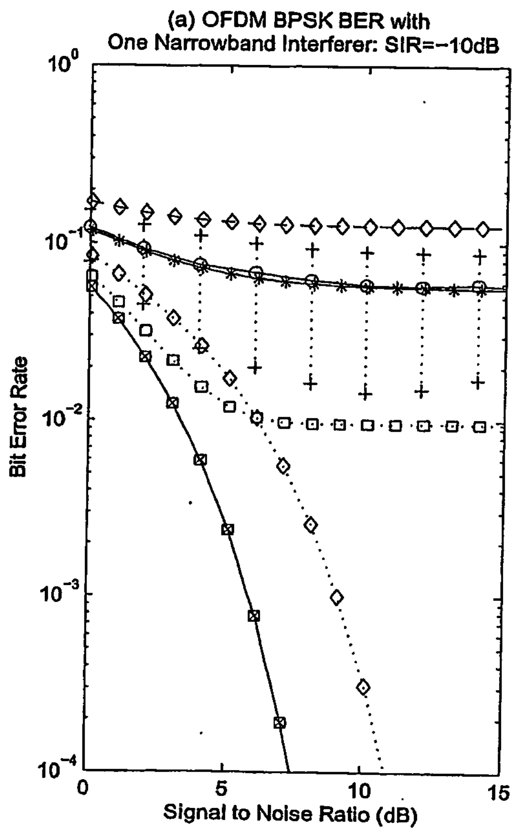


FIGURE 1A

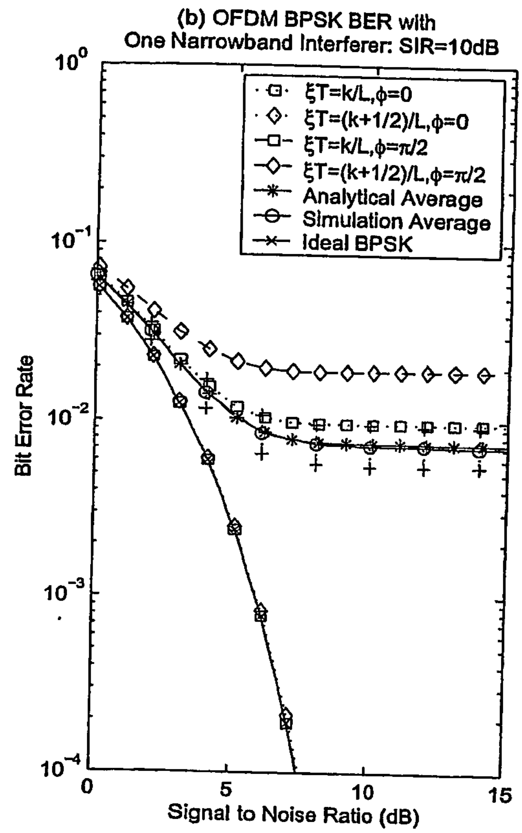


FIGURE 1B

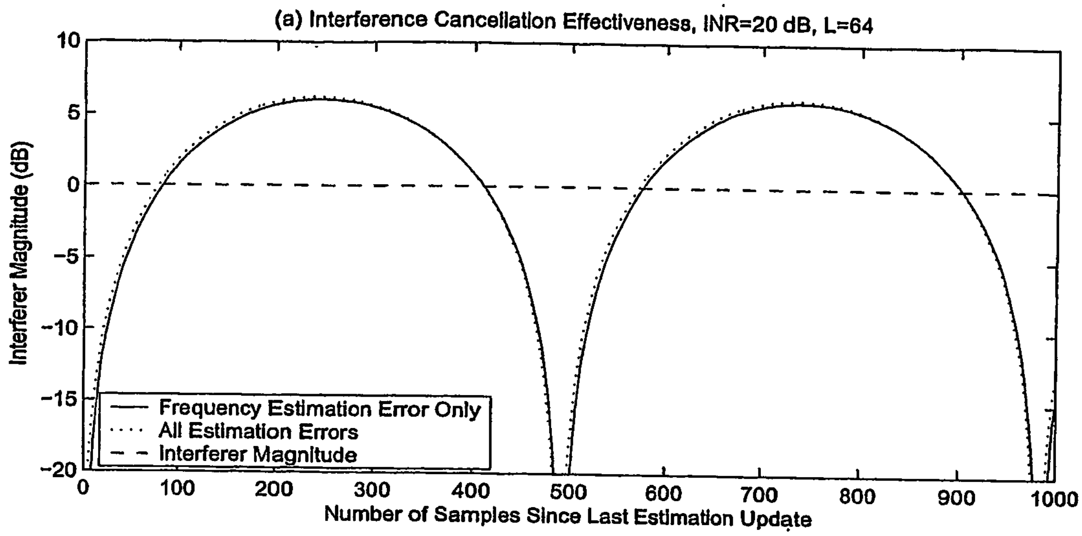


FIGURE 2A

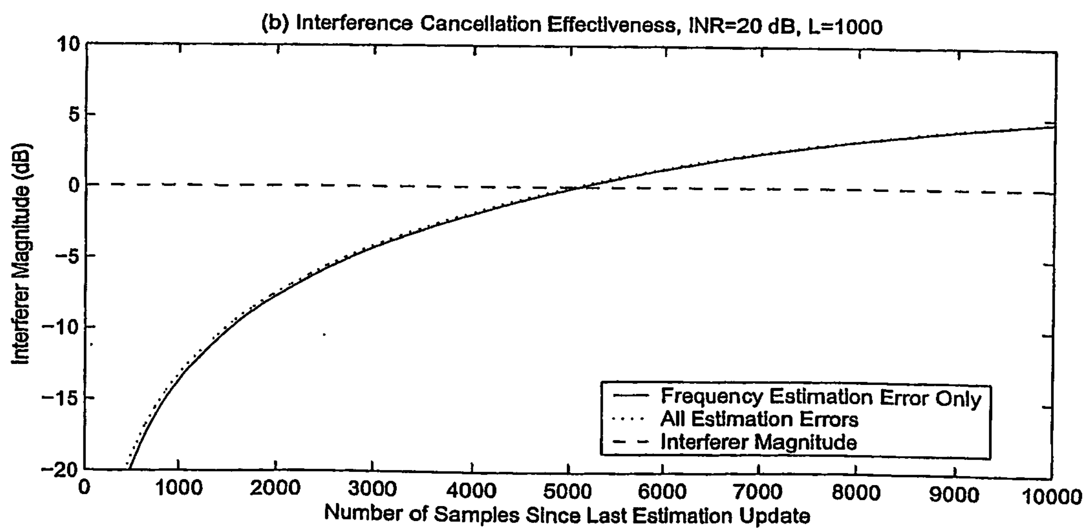
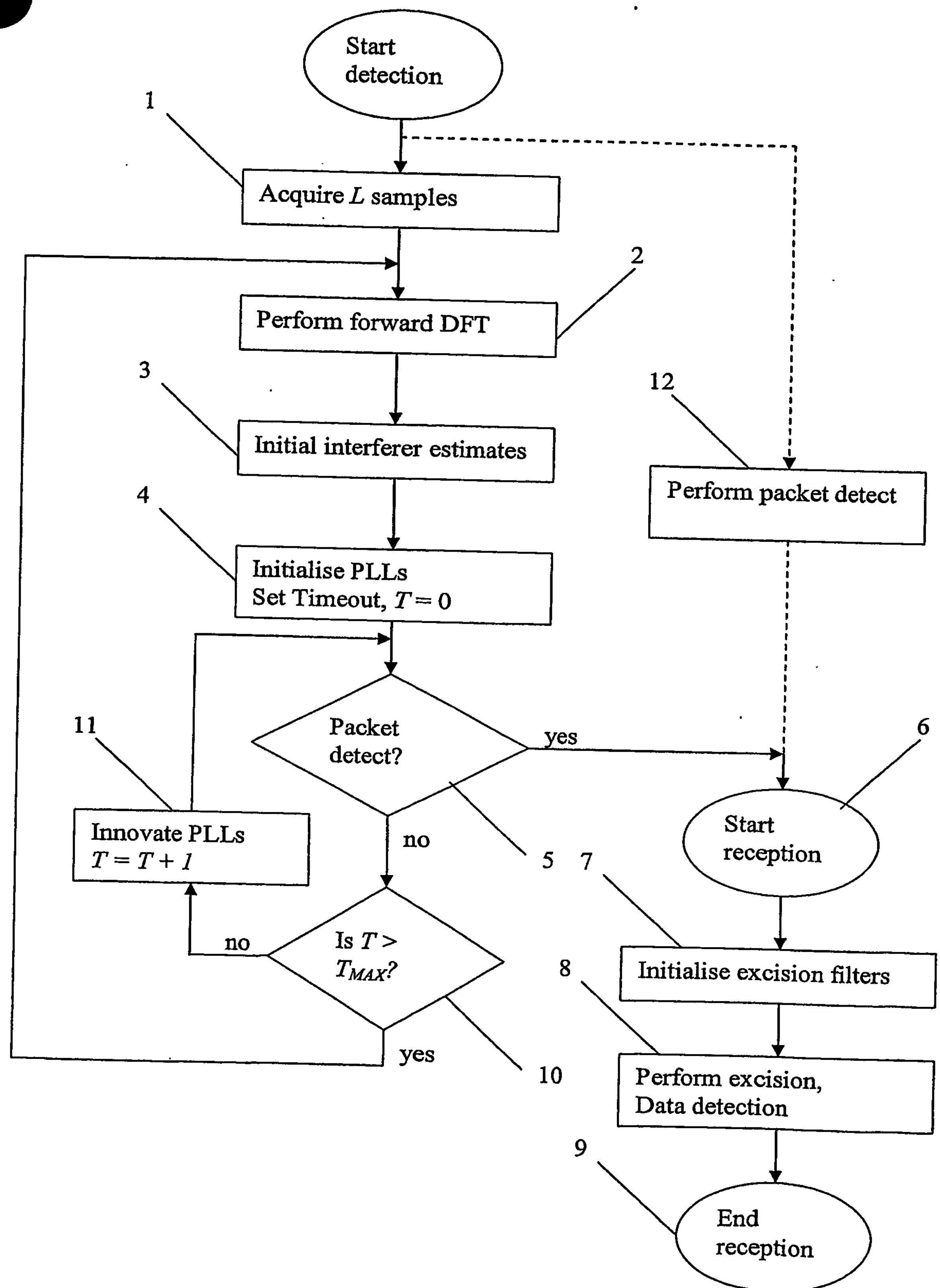


FIGURE 2B



Start detection
1
Acquire L samples
2
Perform forward DFT
3
Initial interferer estimates
12
Perform packet detect
4
Initialise PLLs
Set Timeout, $T = 0$
11
Packet detect?
yes
6
Innovate PLLs
$T = T + 1$
no
5
7
Start reception
no
Is $T > T_{MAX}$?
8
Initialise excision filters
yes
10
Perform excision,
Data detection
9
End reception

Figure 3A

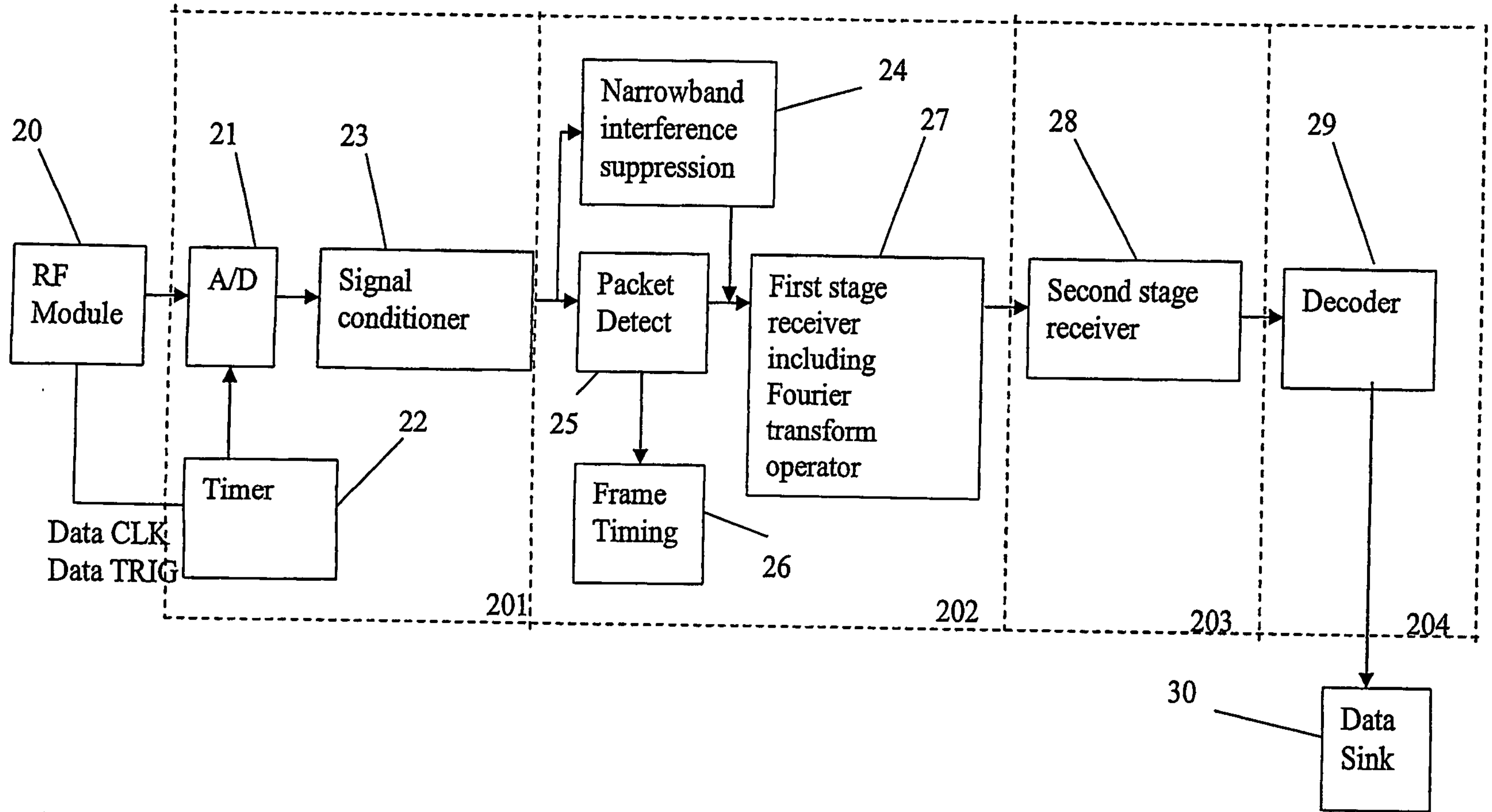

20
RF Module
21
A/D
23
Signal conditioner
22
Timer
Data CLK
Data TRIG
201
24
Narrowband interference suppression
Packet Detect
25
Frame Timing
26
27
First stage receiver including Fourier transform operator
202
28
Second stage receiver
203
29
Decoder
204
30
Data Sink

Figure 3B

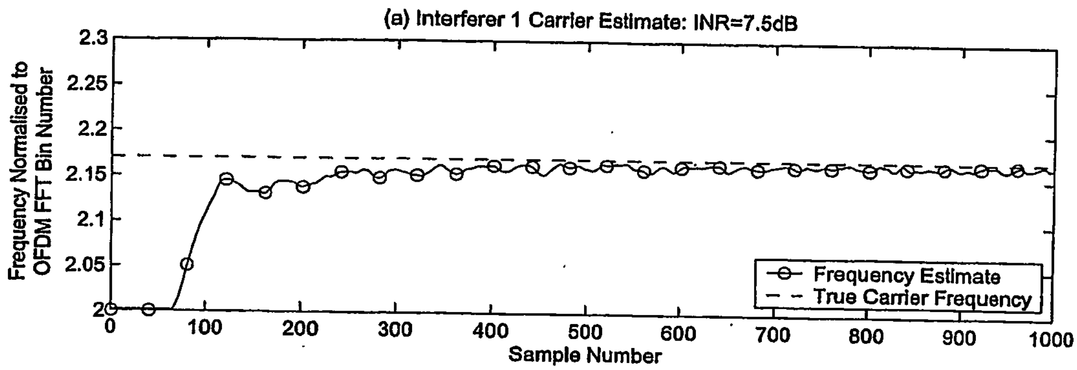


FIGURE 4A

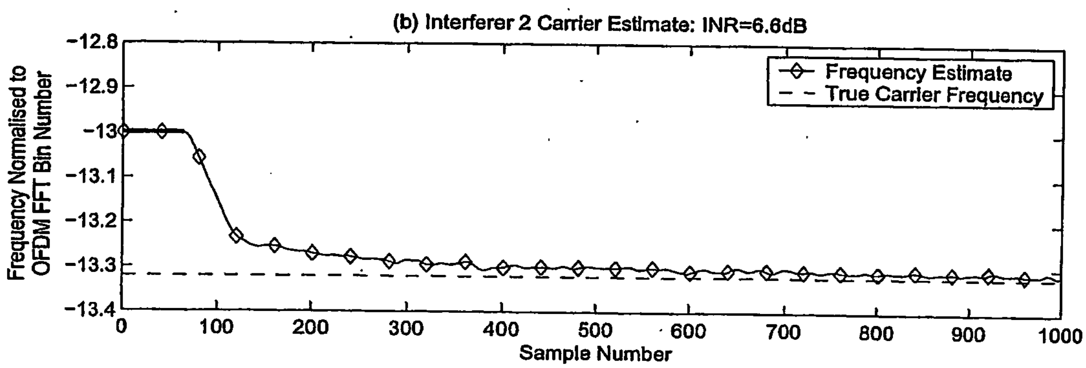


FIGURE 4B

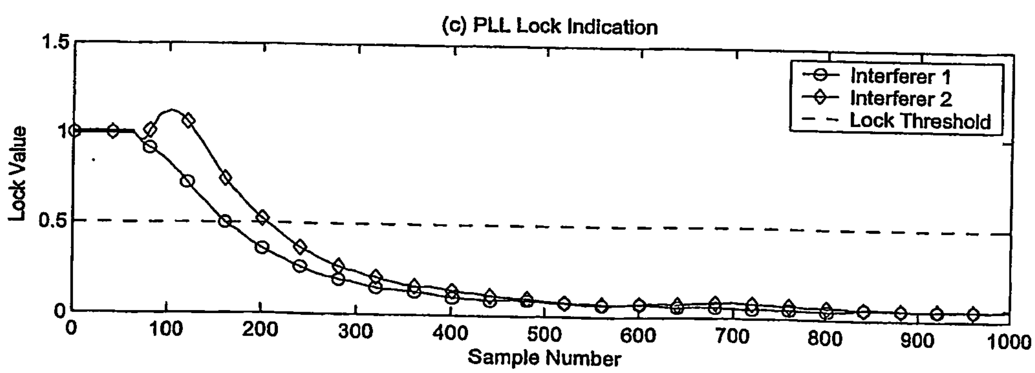


FIGURE 4C

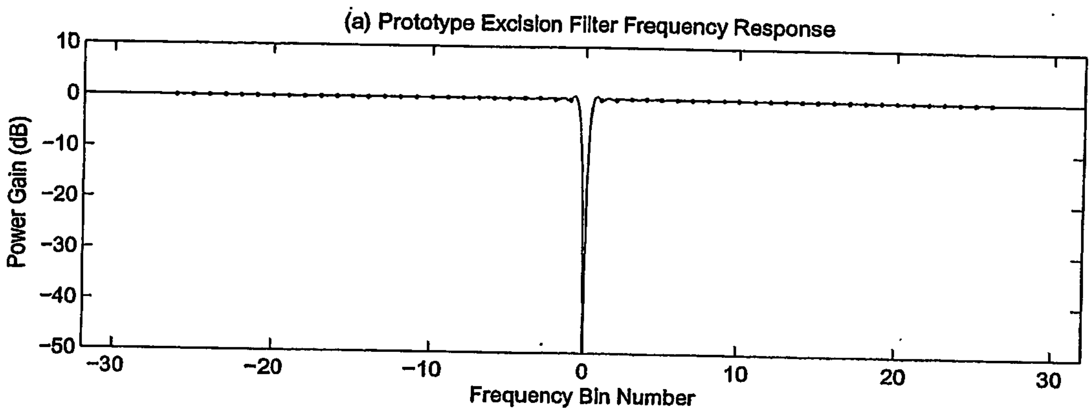


FIGURE 5A

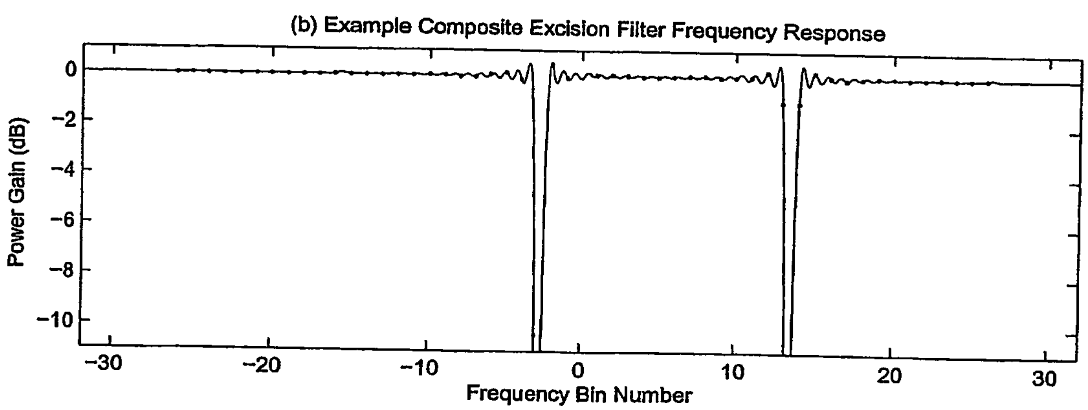


FIGURE 5B

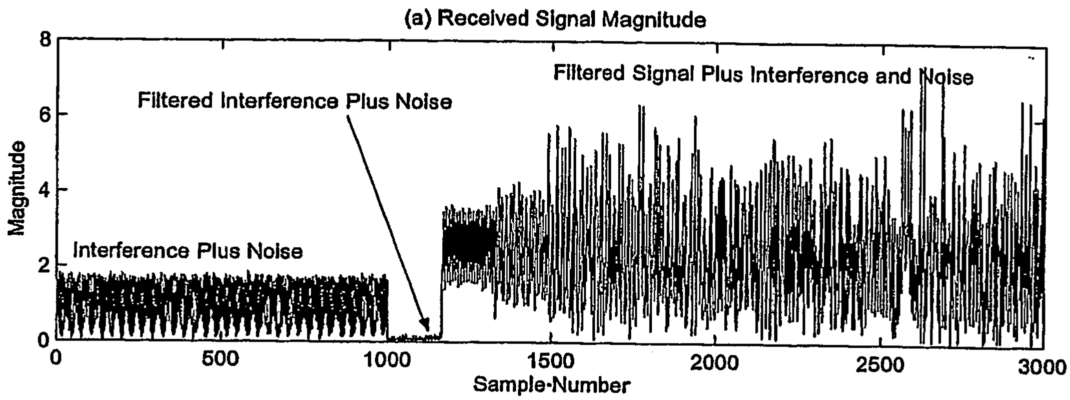


FIGURE 6A

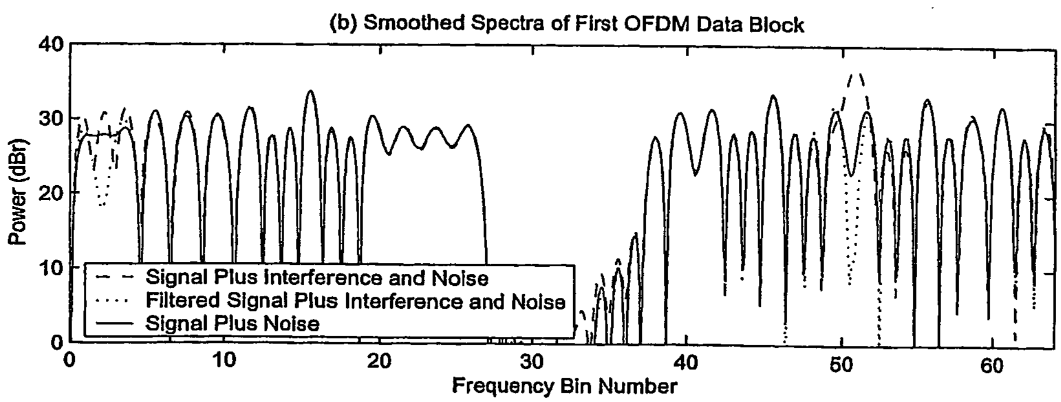


FIGURE 6B

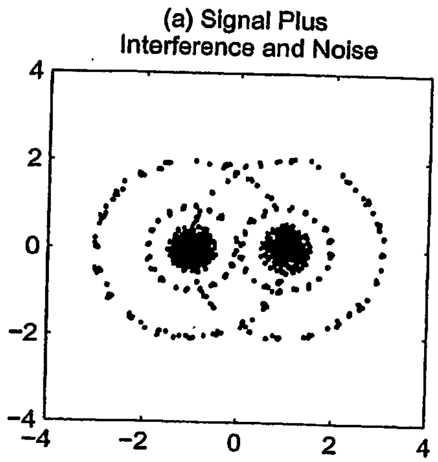


FIGURE 7A

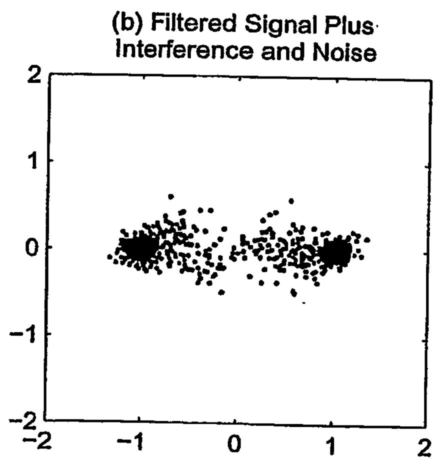


FIGURE 7B

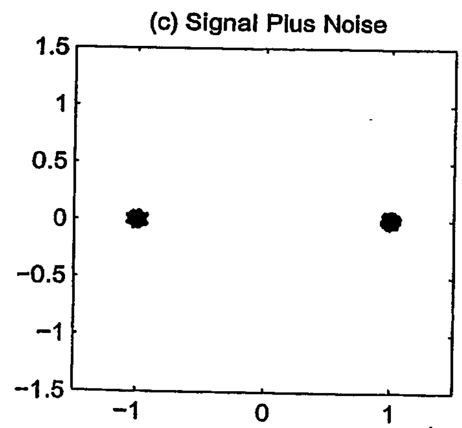


FIGURE 7C

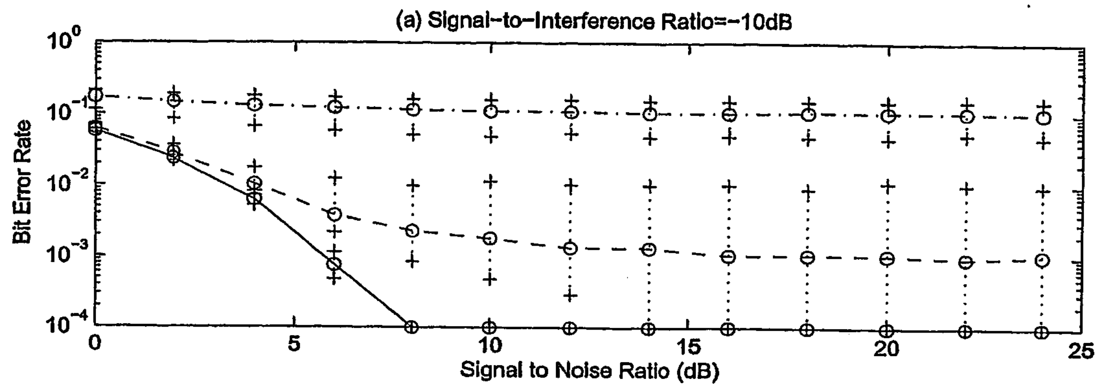


FIGURE 8A

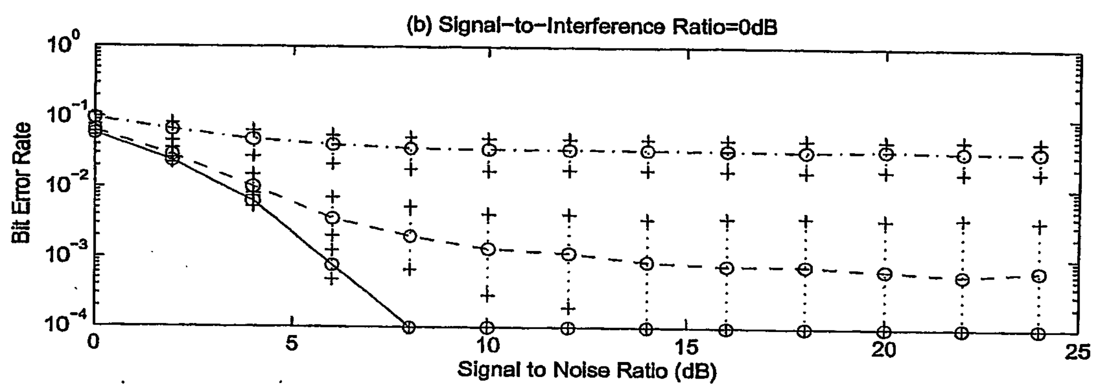


FIGURE 8B

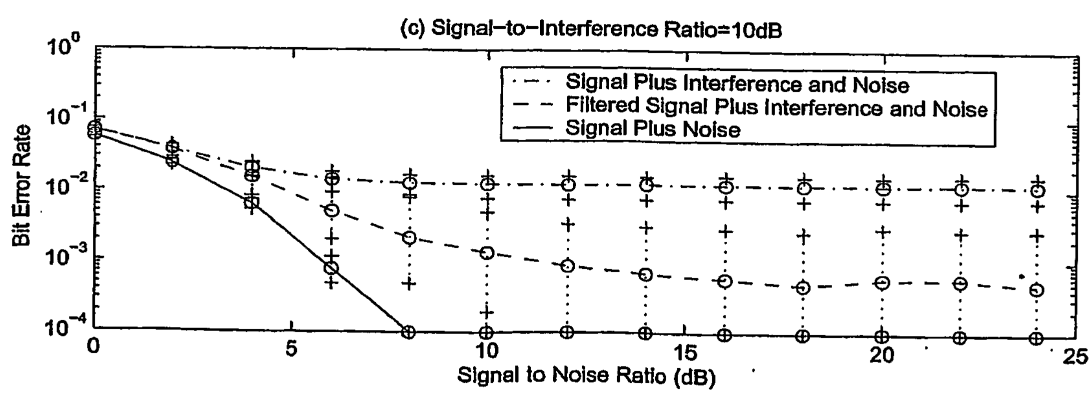


FIGURE 8C

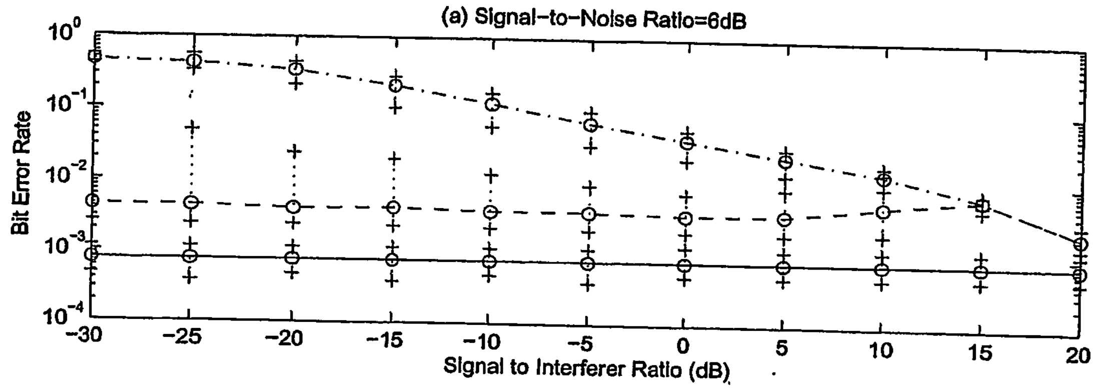


FIGURE 9A

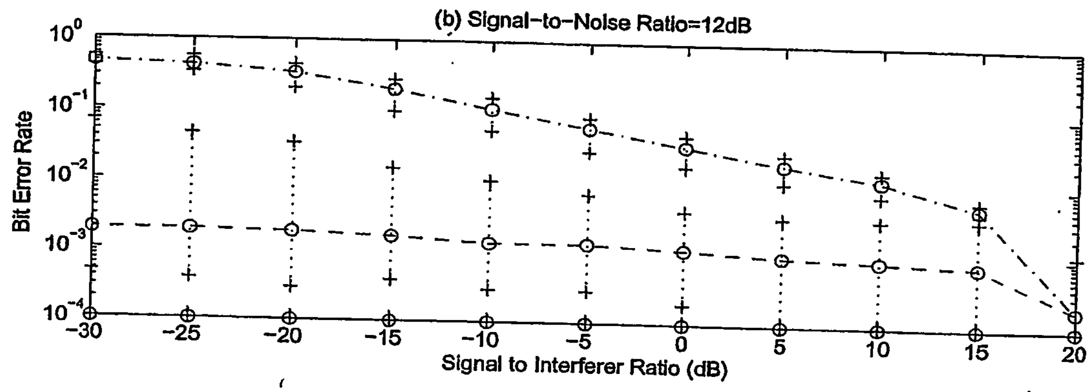


FIGURE 9B

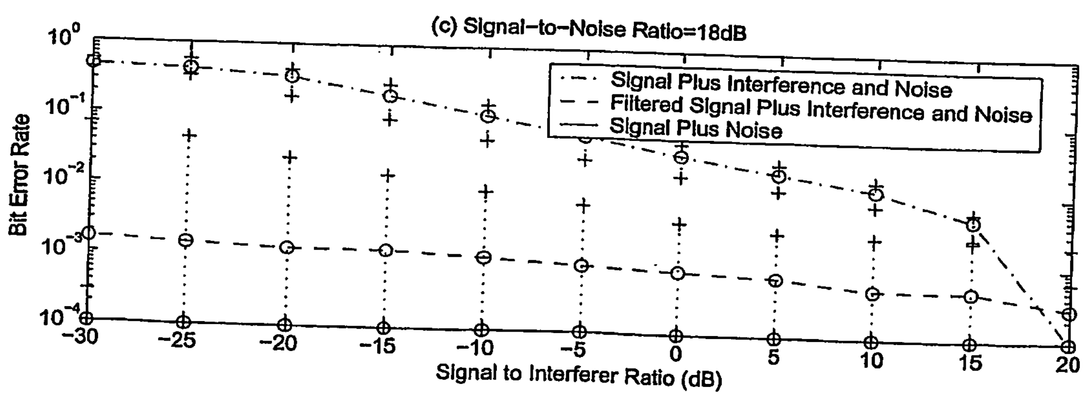


FIGURE 9C

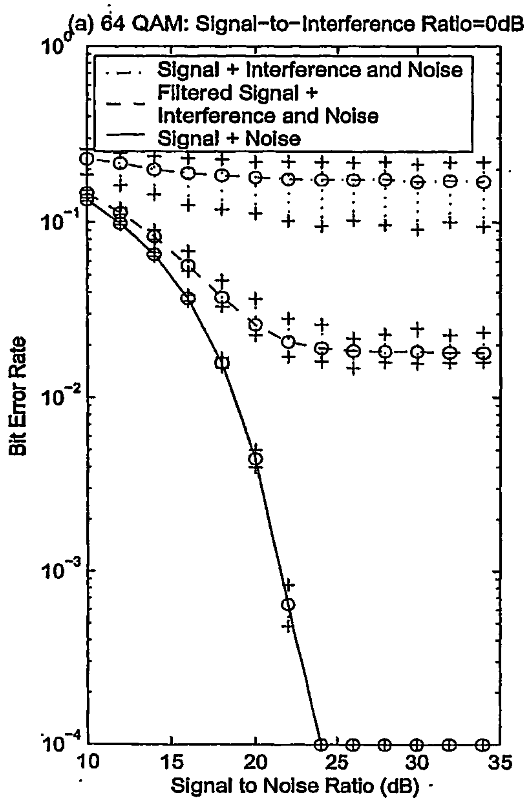


FIGURE 10A

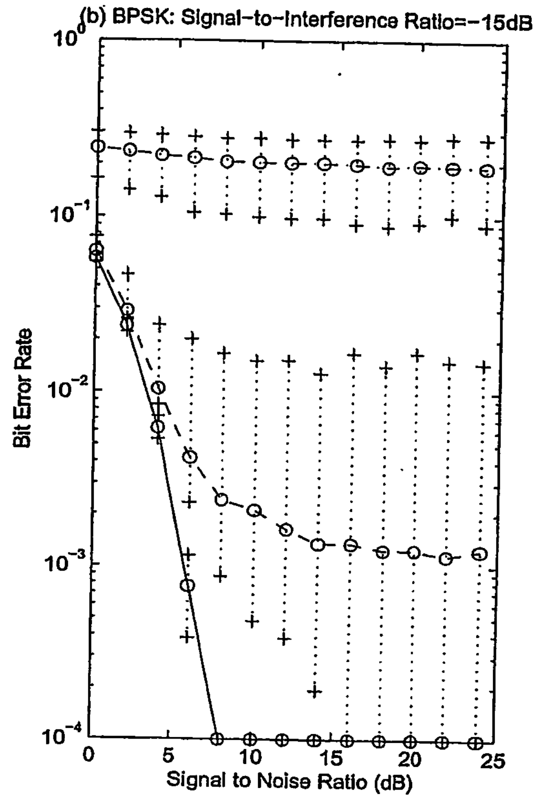


FIGURE 10B